LINEAR INTEGRATED CIRCUITS IN A CHANNELIZED HIGH-FREQUENCY RECEIVER

by

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THESIS

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Linear Integrated Circuits in a Channelized High-Frequency Receiver

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ABSTRACT

An application of modern linear integrated circuits to the design of a channelized high-frequency radio receiver is presented. The results of testing a prototype, crystal-controlled, fixed-tuned integrated-circuit receiver front end and wideband frequency-conversion stage are shown. Problems of image rejection with single frequency conversion are discussed. The design of a channelized receiver section using modern special-purpose linear integrated circuits is shown.

TABLE OF CONTENTS

I. INTRODUCTION

Channelized radio receivers are often used where fixed frequencies or frequency bands are employed. There are many applications which do not require the complete flexibility of a continuous tuned receiver. The speed and precision of channel-switching offered by a flexible, channelized receiver would greatly enhance operations in these applications. The channelized receiver eliminates the need for manual tuning and offers precisely controlled frequency selection when a crystal-controlled local oscillator is used. The use of linear integrated circuits in such a receiver offers a significant reduction in component quantity, greater reliability and lower production costs. With interchangeable or switchable "front end" circuits, receiver flexibility can be greatly enhanced.

The advent of inexpensive, linear integrated circuits has made available a wide variety of circuits for both general and special-purpose usage. Linear integrated circuits are now being used in commercial radio and television receivers for RF amplification, frequency conversion, IF amplification, detection and most recently, low-power audio amplification.

Since linear integrated circuitry enables higher component density than heretofore obtainable, it is possible to construct a channelized receiver system to cover any desired portion of the radio-frequency spectrum using switchable receiver front-end cards with a common IF amplifier and other remaining circuitry. Even with the large number of channels that might be required in order to cover a wide frequency range, equipment size can be kept small by using integrated circuits.

There are many applications of such a fixed-tuned, channelized receiver system, mobile equipment being only one. In fact any system requiring precision tuning, high reliability and high component density resulting in low overall equipment bulk, is a suitable candidate. There are applications in which a wide band (60Hz) intermediate-frequency output is necessary for particular signal-processing requirements. The wideband case was chosen for this study with these applications in mind.

The objective of this investigation is to design and test a prototype fixed-tunedRF amplifier and frequency-conversion stage using linear integrated ciruits and applicable to any portion of the high-frequency radio spectrum (2.0 to 30 MHz). Since the higher frequencies present more design problems, 30.0 MHz was chosen as the operating frequency of the prototype. An IF bandwidth of 45 KHz was selected vice 60 KHz because of the availability of an off-the-shelf, relatively low-cost ceramic filter.

Single-stage frequency conversion was chosen in order to reduce circuit complexity. As described in sections III and IV, the combination of the low input impedance of the transistor amplifier stage and the use of single frequency conversion (30MHz to 455KHz) created some expected problems in obtaining a desirable image-rejection response. Some solutions to the low input impedance problem are given in section IV. Single frequency conversion is desirable for many reasons. In addition to simplicity and reduced circuit complexity resulting in more compact, lower-cost equipment, there are fewer potential spurious responses. These factors outweigh the difficulty caused by single frequency conversion.

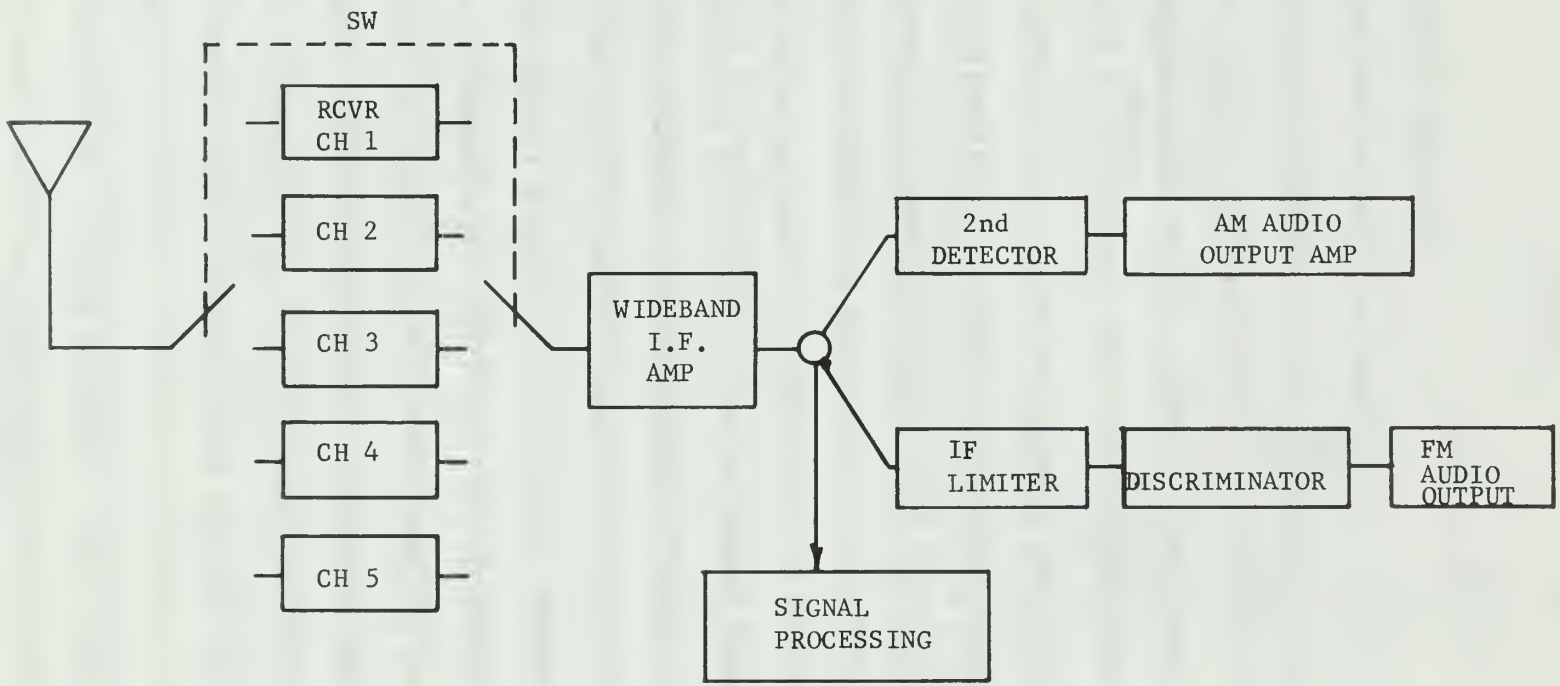


FIGURE 1. Block diagram of a channelized receiver system

The combination of a desired number of fixed-tuned RF/IF sections, a switching system and common IF amplifier, detection and audio amplification sections constitute a simplified, channelized HF receiver system. A block diagram of such a system is shown in Figure 1.

Because of their simplicity and low cost, RCA general-purpose linear integrated circuits were chosen for experimentation. Although external biasing networks had to be designed and included in the circuits, the ease of design change justified this approach.

It is economically feasible to use integrated circuits in a receiver only if a majority of non-active elements can be integrated into a monolithic circuit. Therefore, although the prototype design using general-purpose linear integrated circuits proved the feasibility of the designed circuits, actual production equipment should follow the design of Figure 18. To as large an extent as possible the biasing and other non-active elements should be integrated into a monolithic integrated circuit to provide better system reliability, performance and reduced construction cost.

A digitally controlled, solid-state switching system for channel selection is envisioned. This would provide rapid access to any of the receiver cards, greatly enhancing the flexibility of the system. This is a highly desirable feature in a system with a large number of channels. It is also quite possible to add on IF amplification and detection stages so as to provide simultaneous multichannel operation.

II. CIRCUIT DESIGN

There is a wide variety of special and general-purpose linear integrated circuit configurations available and applicable to HF receiver design. In order to facilitate experimentation, a general purpose IC was sought which would meet the necessary requirements of frequency response, low power consumption, flexibility of interconnection and relatively low cost: The RCA CA 3018A and CA3026 linear integrated circuits were selected. The following paragraphs will discuss some of the more important design criteria for an RF-amplification stage and a frequency-conversion stage. These circuits are intended as laboratory test circuits and do not constitute the entire circuitry of the first sections of a receiver.

A. RF AMPLIFIER

The CA3018A linear IC array, chosen for a single-stage, tuned-RF amplifier is shown schematically below in Figure 2. It is available in a 12 pin TO-5 package.

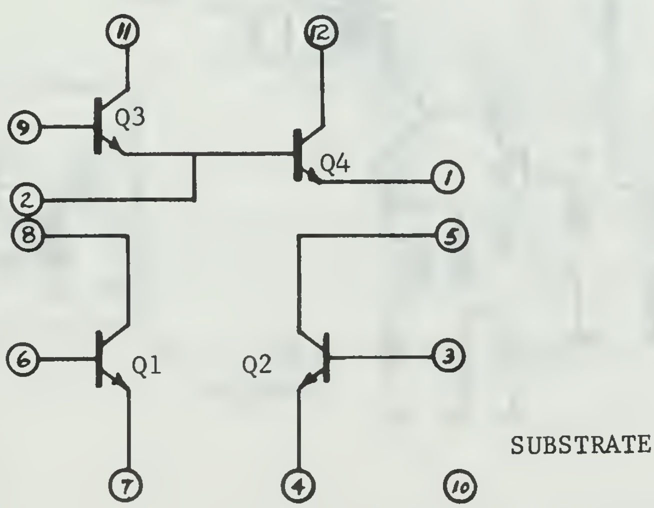


FIGURE 2. The CA3018A Linear IC Array

This configuration lends itself to different connection schemes for experimentation. It was decided that some means of temperature compensation, input signal overload protection and AGC should be provided if possible in the first RF amplifier stage. (As used with a wideband IF stage, AGC is not desirable since a strong but unwanted signal would, through the AGC action, override a less strong but desired nearby signal.) The linearity of an amplifier is therefore extremely important in keeping cross-modulation to a minimum. In this respect a field-effect transistor (FET) would be a better choice for the first RF amplifier stage. As metal-oxide semi-conductor technology progresses it will be possible to obtain low-cost monolithic integrated circuits containing FET's.

The tuned-circuit design of Figure 3 provides the desired circuit characteristics. It was adapted from a similar circuit described in Ref. 6.

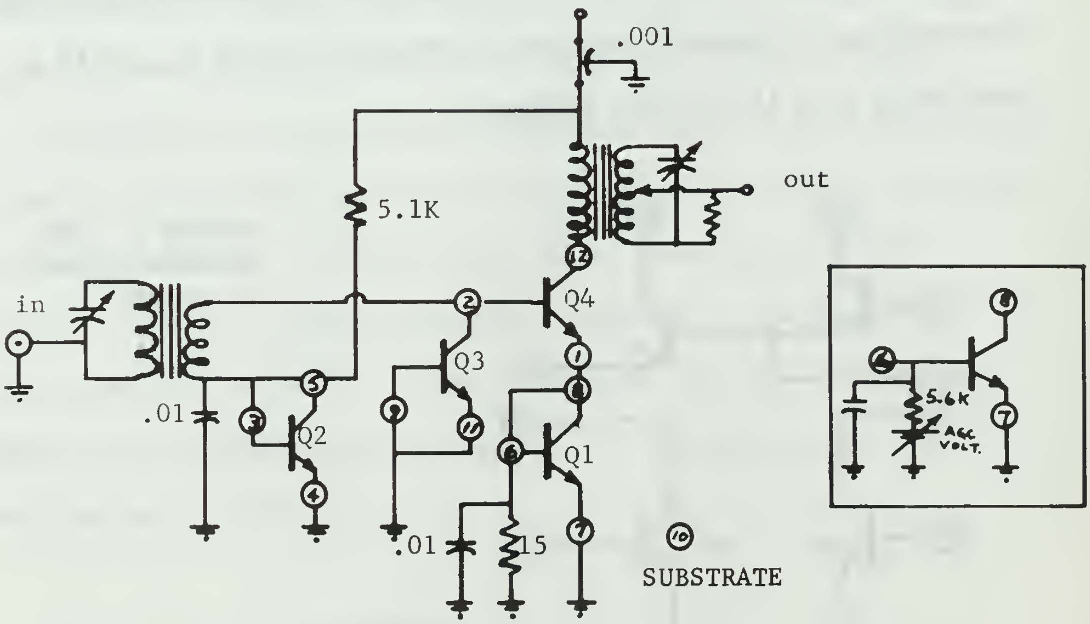


FIGURE 3. RF Amplifier Stage

The inset shows the connections to Q2 which provide AGC when required. Here Q2 is normally saturated. A negative-going signal at the base resistor brings Q2 out of saturation greatly increasing the emitter resistance of Q4 and thus provides a very effective automatic gain control.

Transistor Q1 is connected so as to use the base-emitter junction as a forward-biased diode holding the base of Q4 at a DC level which changes with temperature as Q4 changes. Thus temperature compensation is obtained.

Transistor Q3 is used as a reverse-biased diode to provide signal overload protection at the input. This protection is required when the receiver is located near a transmitter as in a commercial or military channelized transceiver.

1. Stability

Amplifier stability may be investigated using two-port network y-parameter analysis. The amplifier section is considered as a two-port network and the stability criterion used here is a function of the y-parameters of the two-port network [Reference (4)]. The stability factor is given by:

$$C = \frac{y_{12}y_{21}}{2g_{11}g_{22} - \mathcal{Re}(y_{12}y_{21})} .$$

For the range $0 \leq C \leq 1$ the network is inherently stable.

Since transistors Q1, Q2 and Q3 in this circuit can be ignored in a small-signal equivalent circuit, the necessary y-parameters are those of the transistor Q4 alone. These may be found from the specifications sheets and at 30MHz are:

$$y_{ie} = 0.75 + j1.2 \text{ mmho} \qquad y_{fe} = 23 - j13 \text{ mmho}$$

$$y_{re} = 0 - j0.1 \quad '' \qquad y_{oe} = 0.75 + j0.2 \quad ''$$

From these one can evaluate C;

$$C = \frac{(0.1)(23^2+13^2)^{\frac{1}{2}}}{(2)(.75)(.75)-(-1.3)} = 1.035 \; .$$

Thus the transistor is potentially unstable at 30MHz; that is, there are some source and load admittances which will allow the circuit to oscillate. If the source and load admittances are included, it is required that:

$$\left\{2(g_{ie}+G_s)(g_{oe}+G_L) - \mathcal{R}e\left[y_{fe}y_{re}\right] - \left|y_{fe}y_{re}\right|\right\} > 0 \; .$$

Reference 11 refers.

It is thus clear that the stability of an amplifier may be controlled by the selection of G_s and G_L or by the transformation ratios of the transformers used to couple the input and output loads to the amplifier.

Assuming that the transformation ratios are such that $G_s = G_L = 1$ mmho it follows that:

$$\left[2(.75+1)(.75+1)-(-1.3)-2.64\right] = 4.78 \; .$$

Thus the amplifier is stable for these values of source and load admittance.

The choice of 1 m℧ for G_s and G_l while arbitrary, is realistic. The tuned circuits at the input and output must have a very high Q in order to provide the necessary selectivity for image rejection. By using the proper turns ratios the very high impedance of the tuned circuits may be transformed to the desired 1KΩ level (G = 1 mmho) without excessively loading the tuned circuits.

2. <u>Gain</u>

The voltage gain of the RF amplifier can be calculated by using a small-signal model of the tuned amplifier. Figure 4 shows such a model for a common-emitter configuration.

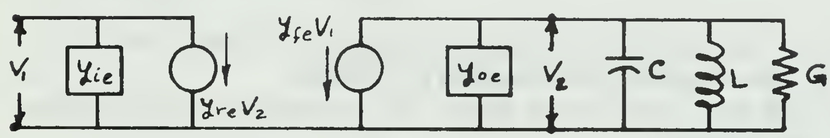


FIGURE 4. Small signal model for a tuned amplifier.

The gain is

$$A_v = V_2/V_1 = y_{fe}/(y_{oe} + j\omega C + G + \frac{1}{j\omega L})$$

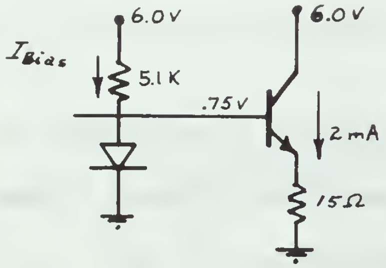
At resonance the impedance of the tank becomes resistive and

$$|Av(\omega o)| \equiv Ao = \frac{|y_{fe}|}{g_{oe} + G}$$

For stability, the load reflected into the amplifier by suitable impedance transformation was chosen to be 1 m ; therefore

$$Ao = \frac{26.4}{.75 + 1} = 15 \quad .$$

Biasing of the amplifier is accomplished by using Q1 to set the voltage at the base or Q4. In the configuration in which the AGC is by-passed, the emitter resistor (15Ω) limits the emitter current of Q4 to the proper level. With an emitter current of two mA, $V_{BE} = 0.72v$; then:



$V_B = .75V \quad V_E = 0.03V$

$I_B = I_E/h_{fe} = 0.018$ mA

and

$$I_{BIAS} = \frac{5.25}{5.1K} = 1.03 \text{ mA}$$

The total DC drain current is therefore 3.03mA.

3. Image Rejection

The image-rejection performance of a superhetrodyne receiver is principally a function of two factors: (a) The difference in frequency between the desired radio frequency and the local oscillator and (b) the Q of the tuned circuits of the RF amplification stage.

In the interests of simplicity, it was desired that single frequency conversion be used to provide an IF of 455KHz. This indicates that the image frequency occurs at 30.910MHz when the receiver is tuned to 30.0MHz. The bandwidth required of the tuned circuits is therefore less than 1.8MHz. Taking a 3-db bandwidth of 227.5KHz the unloaded Q of the tuned circuits is:

$$Qo = fo/BW = \frac{30.00\text{MHz}}{227.5\text{KHz}} = 132$$

This indicates that a signal of 30.91MHz which is 8 "bandwidths" from 30.0 will be down 87% or -27.7db. By using small toroids in the input/output tuned circuits and coupling in the proper loads, the loaded Q of the tuned circuits can be made as high as 150, and thus should provide adequate image rejection for most applications. (A crystal filter would provide greater image rejection and selectively owing to the "steeper skirts" of the response curve. Crystal filters are however much more costly than LC tuned circuits). At lower frequencies the image characteristics would progressively improve.

B. FREQUENCY CONVERSION

The RCA CA3026 integrated circuit offers the capability of combining the functions of local oscillator and mixer in one unit. In the usage presented here the local oscillator is crystal controlled because of the channelized nature of the receiver. It should be noted that in different applications, a tunable local oscillator could easily replace the crystal oscillator without altering the rest of the mixer section. It was decided to use one of the differential amplifiers as an oscillator and the other as a mixer. A modified Pierce circuit was used for the oscillator and is shown in Figure 6. A schematic diagram of the mixer is shown in Figure 7.

1. Local Oscillator

The CA3026 integrated circuit contains two independent differential amplifiers with associated constant-current transistors. A schematic diagram of the CA3026 is shown in Figure 5.

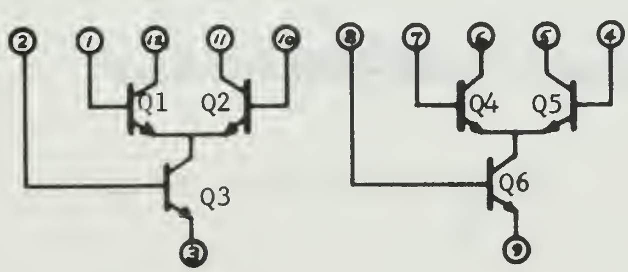


FIGURE 5. Schematic diagram for CA3026

The oscillator is essentially comprised of Q2, the crystal and the associated biasing network. Q3 provides a constant-current source for Q1 and Q2 while Q2 provides buffering for the oscillator.

Conventional biasing is used in the oscillator section. Q1 and Q2 are biased for 1mA collector current with pin 5 at a 4-v dc level. Stability is achieved by maintaining approximately 6 I_B thru, R_3, R_5 and R_7.

After building and testing the oscillator it was found that a supply of 7.5 volts was required vice 6 volts for proper circuit performance. (A somewhat different configuration was later found to work better with this crystal. See appendix A).

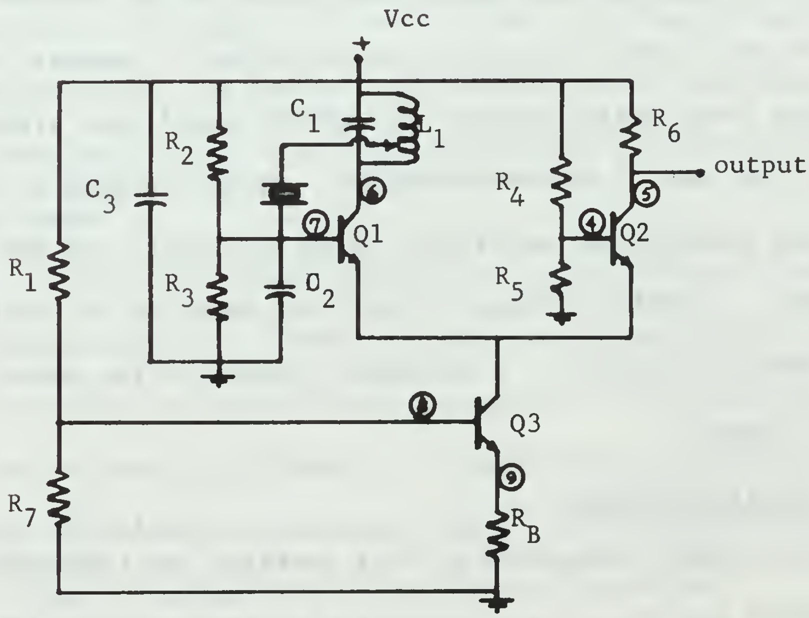


FIGURE 6. Schematic diagram of crystal-controlled local Oscillator.

The crystal used in the experimental circuit is an inexpensive general-purpose crystal (.01% tolerance) designed for use in transistor circuits on a third overtone at 30.455MHz. In the configuration used, the series-resonant mode of the crystal is used to provide a low-impedance signal path from collector to base of Q1.

The tank circuit formed by C_1 and L_1 is tuned to 30.455MHz. Capacitor C_2 (100 pF) is necessary in order to smooth out the waveform. When properly adjusted the output of the oscillator is 250mv peak-to-peak. Figure 10 shows sketches of oscillator waveforms.

a. Oscillator Circuit Biasing

As mentioned earlier the transistors Q1 and Q2 are biased at 1mA. Q3 is therefore biased for 2Ma, and choosing 1.0v at pin 9, RE = 0.5k. The voltage at pin 8 is 1.7v, therefore

$$I_{B3} \doteq \frac{2mA}{h_{fe}} = 0.02mA$$

$$R_7 = 1.7v/GI_{B3} = 14.2K$$

and $$R_1 = (6.0 - 1.7)/7I_{B3} = 37K$$

Assuming a 2-volt drop across Q3, the base voltage of Q1 and Q2 will be 3.7v and the base currents are 0.01mA. This gives

$$R_2 = R_5 = 2.3v/0.07mA = 32.9K$$

$$R_3 = R_5 = 3.7v/0.06mA = 61.7K$$

Q2 is biased to provide a peak output swing of 1.5v so that the collector voltage of Q3 is 4.5v and

$$R_6 = (6.0 - 4.5)/1.0\text{mA} = 1.5\text{K}$$

The oscillator tank circuit is comprised of an inductor wound on a tunable ceramic form and tapped for the crystal connection. The capacitor C_1 is variable 1.5 to 7.0pF.

2. Mixer

The second differential amplifier of the CA3026 integrated circuit, connected as shown in Figure 7, operates as a mixer. It has been shown [Ref.13] that when two signals, relatively close in frequency, are super-imposed the result is an amplitude-modulated signal with an envelope which pulsates in amplitude at the difference frequency. It has also been shown that the amplitude of the envelope can be described by the equation

$$\text{Envelope} = AoEo(1 + M, \text{Cos } wt - m_2\text{Cos } 2wt + m_3 \text{ Cos } 3wt...)$$

where Eo is the amplitude of the larger oscillation, and Ao, M, M2, etc. depend upon the ratio Es/Eo. Es is the signal voltage and Eo is the local oscillator voltage. See Figure 8. The equation for the envelope shows that when one signal is much larger than the other, the envelope is essentially sinusoidal, proportional to the weaker signal and independent of the amplitude of the stronger signal. As the two signals approach the same amplitude, distortion in the envelope increases and reaches a maximum of 20 percent second harmonic when the signals are of equal amplitude [Ref. 14].

It is therefore desirable to ensure that the local oscillator signal is much larger than the incoming signal at the mixer.

a. Circuit Description

As can be seen from Figure 7, the incoming signal is transformed to a push-pull output via the center-tapped input transformer and the transistors Q5 and Q6. The oscillator signal is in push-push through Q5 and Q6 thus it tends to cancel itself out in the output. The input transformer is would on a small toroid and has an unloaded Q of approximately 100. The output transformer is a sub-miniature IF transformer tuned to 455KHz.; it has an unloaded Q of 70.

Since the RF input transformer is center-tapped to ground, a negative voltage (-Vee) is required. The transistors are biased at the same operating currents as the local oscillator section, the biasing of Q4 being exactly analogous to that of Q3. See Figure 6.

A Clevite Ceramic filter with a bandwidth of 45KHz centered at 455KHz follows the IF transformer. The low input impedance of the ceramic filter loads the IF output tank extending the bandwidth to the limits of the filter. A response curve for the Clevite filter is shown in Figure 9.

The conversion gain of the mixer stage is defined as the ratio of the IF output voltage to the signal voltage applied at the input. A conversion gain of the order of unity is typical for this type of circuitry.

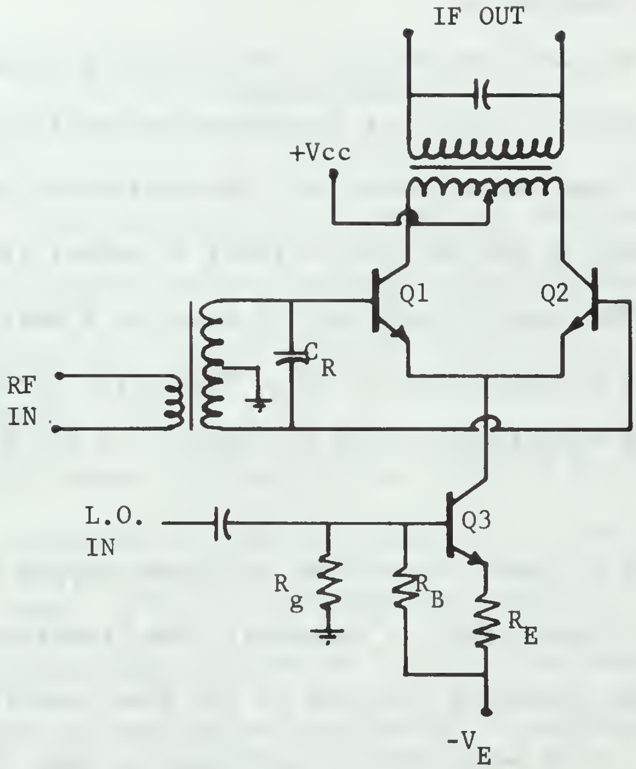


FIGURE 7. Mixer Section

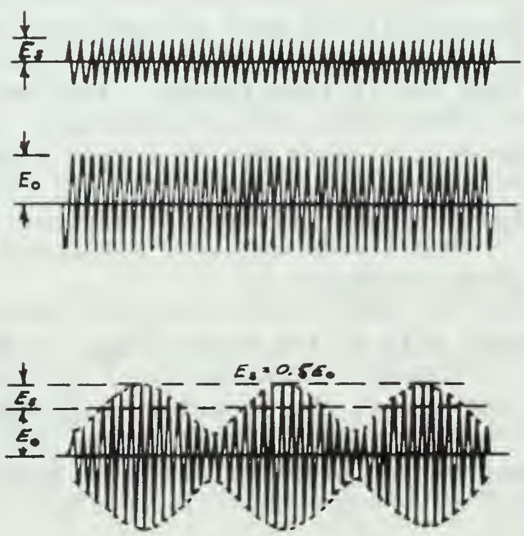


FIGURE 8. Heterdyning two signals of slightly different frequency. The envelope of the resultant is the difference frequency.

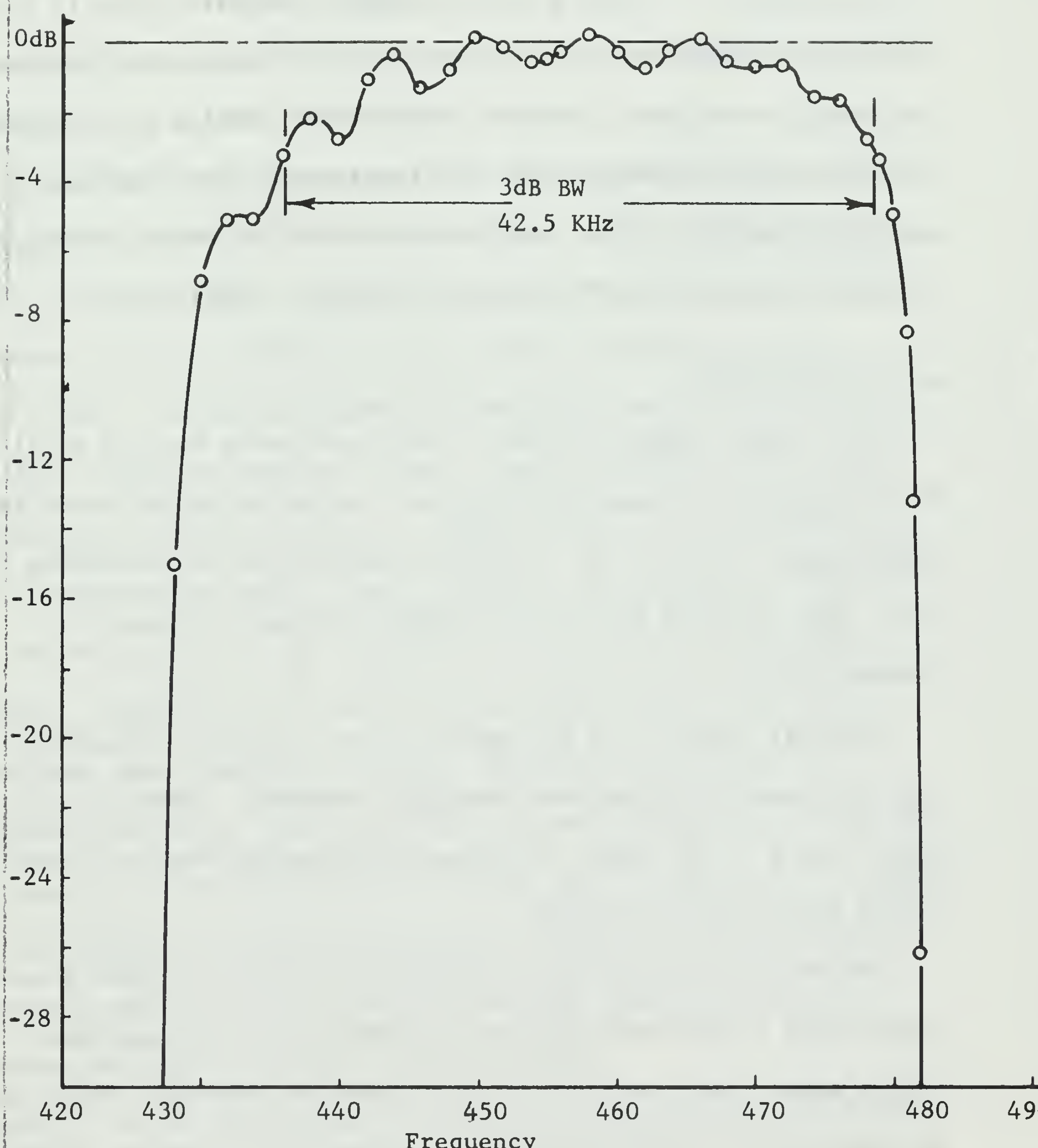


FIGURE 9. Clevite Ceramic Filter Response. Filter No. TL-45D65A, Nominal 6dB BW = 45 KHz. Insertion loss @ 27 C = 4dB Shape factor = 1.4:1

III. EXPERIMENTAL RESULTS

The RF amplifier and local oscillator/mixer circuits discussed in Section II were constructed and tested in order to demonstrate the feasibility of using general-purpose integrated circuits in a channelized 30MHz receiver. Except as noted below, the circuits performed as designed. The most significant problem is the result of the low input impedance of the transistors. This can cause excessive loading of the tuned circuits reducing the Q of these circuits below the level required for proper selectivity.

A. RF AMPLIFIER

The circuit shown in Figure 3 was constructed using a small micarta board as a mounting platform. Isolation of the input and output tuned circuits was obtained by sectioning the enclosing box. This is shown in the photographs of Figures 20 and 21 (Appendix B).

As first constructed the amplifier was unstable indicating that the tuned circuits were improperly matched. Increasing the turns ratios at the input and output and tapping down the input solved the stability problem.

The measured voltage gain of the amplifier is 10 (20db power gain) which is 2db less than the calculated gain. Measurement errors might easily account for this small difference. There was no distortion of the signal over the input level range of 50 μV - 100 mV. A modulated signal (400Hz and 100 Hz) was amplified by the test circuit, connected to an antenna and radiated. When it was picked up by a commercial receiver and detected, no distortion in the modulation was observed.

The dynamic range of the amplifier expressed in decibels can be defined as

$$10 \text{ Log}_{10} \left(\frac{Vmax}{Vmin}\right)$$

where Vmax is the maximum input signal for an undistorted output and Vmin is the minimum input signal level.

The maximum and minimum available undistorted RF signal levels (30MHz) were 100 mV and 50 μV respectively. These values yield a dynamic range of 33db. It sould be noted that there is no reason to suspect that the amplifier is not linear for signal levels below 50 μV. At the same time the amplifier should remain linear for input levels as high as 1 volt. Thus the dynamic range figure of 33db is considered to be very conservative; 53db is probably a more accurate value with a lower limit set by the noise threshold.

The selectivity of the amplifier stage was much lower than desired. Initially the loading of the tank circuit by the low transistor input impedance caused a decrease in Q so that the bandwidth was of the order of 10MHz. By decreasing the loading and yet maintaining stability, the Q of the tuned circuits was increased: A 3-db bandwidth of 2MHz was obtained. This is still not adequate to provide image rejection and further work needs to be done to increase the loaded Q of the tuned circuits.

The total DC power drain for the amplifier stage was measured at 6.05 milliwatts (1.01mA at 6 volts).

A temperature stability evaluation has not been conducted; however the technique of using a diode in the base-biasing circuit is well known and is highly effective.

The noise figure of the amplifier is estimated from the data provided by the IC manufacturer to be approximately 5db as a maximum value.

B. FREQUENCY CONVERSION

The local oscillator and mixer circuit shown in Figures 6 and 7 were first built as a breadboard model. The oscillator exhibited a definite tendency toward the fifth overtone vice the desired third overtone oscillation and the mixer circuit using a toroidal transformer in the output could not be easily tuned to 455 KHz.

In order to overcome these problems, a printed circuit board was made and the circuits re-built. A sub-miniature, tunable IF transformer was substituted for the output toroid of the mixer section. With the oscillator and mixer sections on the printed circuit board the bias network components could be conveniently changed as required to establish proper bias levels. (This was necessary because it was found that the supply voltage had to be raised to 7.5V to drive the oscillator crystal). The addition of C_2 smoothed out the waveform and helped to eliminate the fifth overtone oscillations. Proper tuning of L_1 and C_1 was somewhat critical but when achieved eliminated the unwanted higher-order modes of oscillation and maximized the output at 250mV. Figures 22 through 24 (Appendix B) show the prototype circuit.

The oscillator output is nearly sinusoidal as shown in Figure 10. The output of the mixer reflects the non-sinusoidal characteristics of the oscillator signal and only generally resembles the ideal case of Figure 8, as is shown in Figure 11, since the output contains not only the difference frequency but all of the other frequencies which appear in the equation for the envelope as well. Unfortunately waveform photographs could not be obtained because the oscilloscope used was not adaptable to the available cameras.

With the RF input at 50mV the 455-KHz component of the output was observed to be 50mV (by using a wave analizer); thus the conversion gain of the mixer is unity as expected.

The bandwidth of the unloaded IF output transformer is 6.5KHz and is therefore much too narrow for use in a wideband IF amplifier design; its use here, allowed as simple means of determining the characteristics of the mixer stage.

By using the low input impedance of a ceramic filter to load the tuned output tank the bandwidth is extended to the limits of the ceramic filter. At the same time there is an accompanying decrease in signal level. Receiver selectivity is essentially determined by the ceramic filter.

A schematic diagram illustrating the use of the ceramic filter in conjunction with a CA3018A integrated circuit used as an IF amplifier is shown in Figure 12. Although this circuit was not actually constructed, experimentation with a similar circuit using discrete components demonstrated the capabilities of the feedback amplifier shown. This and other proposed circuits are discussed in the following section.

The results of testing the RF amplifier and frequency-conversion circuits are summarized in Table 1.

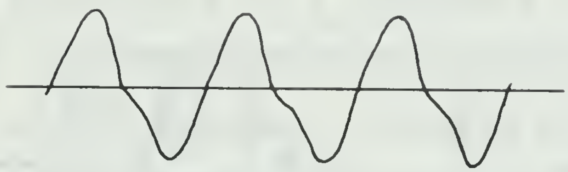


FIGURE 10. Local oscillator signal

FIGURE 11. Mixer output 50mV at 455KHz

TABLE I

Performance Characteristics

RF Amplifier		Mixer	
Power Gain	20 db	Local Oscillator Frequency	
Bandwidth	2 MHz	30.455MHz $\pm$ 0.01%	
Noise figure	5 db	Oscillator Output	250mV
Dynamic Range	33 db	Mixer output	455KHz
		Mixer Gain	1

IV. CONCLUSION

The circuits of Figures 2, 5, and 6 demonstrate in simplified terms, a design approach for the first stages of a channelized HF receiver. Some of the more recent techniques for biasing, circuit stablization and multiple function applications have been studied.

For reasons of economy, simplicity and ease of demonstration, general-purpose integrated circuits have been used rather than the more complex special designs which are now becoming available. Photographs of the prototype circuits are presented in Appendix B.

The problem of the low input impedance of the RF amplifier can be solved by any of several well known techniques. The most common method is probably the use of an emitter-follower in the input. An extension of this approach leads to the Darlington pair shown in Figure 12. This is facilitated by the design of the CA3018A in that Q3 and Q4 form the necessary cascaded common-collector pair. The bootstrp capacitors C_1 and C_2 greatly increase the input resistance of the pair and thus decrease the loading of the input tank. The input resistance for the circuit shown can be calculated from [Reference 5] :

$$R_i = \frac{R_{eff} h_{fe1} h_{fe2} Re}{R_{eff} + h_{fe1} h_{fe2} Re}$$

where $R_{eff} = \dfrac{R_3}{1 - Av}$ and Av is the voltage gain of the pair.

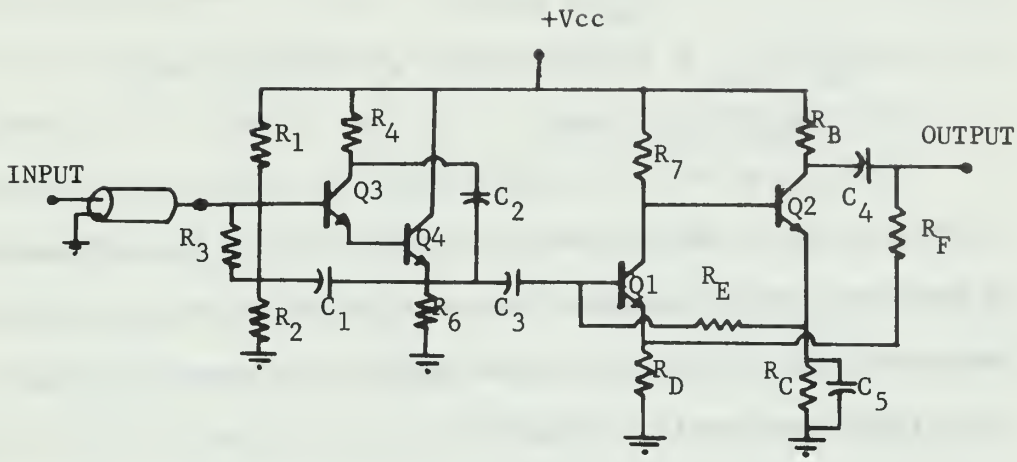


FIGURE 12. IF Amplifier using a ceramic filter. A bootstrapped Darlington Pair is followed by a negative-feedback amplifier.

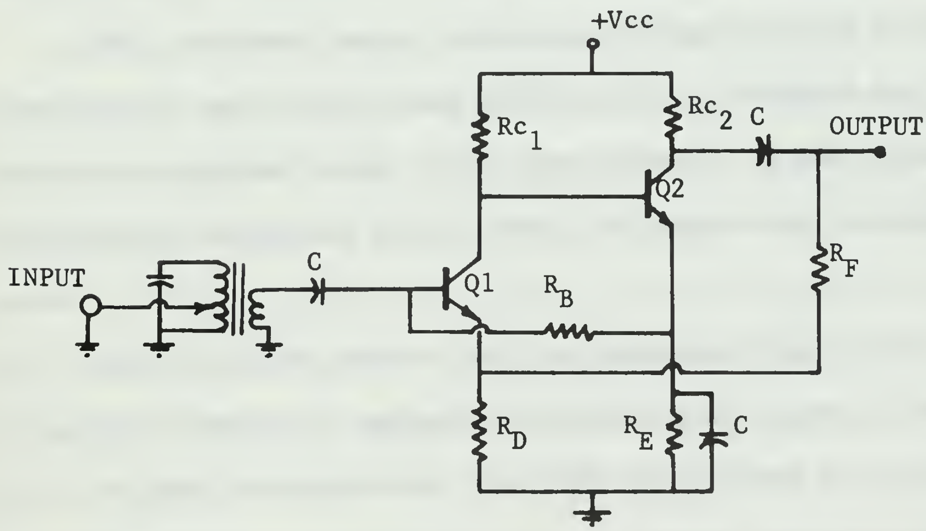


FIGURE 13. RF Amplifier using negative feedback to provide high input and low output impedance.

Typically Av = 0.95. For Re = 4K and R_3 = 100K (typical values):

$$R_{eff} = 2\ M\Omega$$

$h_{fe1} - h_{fe2} = 110$, therefore $R_i' = 48\ M\Omega$ where $R_i' = h_{fe1}h_{fe2}Re$, then

$$R_i = 2\ M\Omega$$

This value of input impedance is sufficient to prevent loading of the input tuned circuits. One other desirable feature of the Darlington pair is that the output impedance is lower than that of a single-stage emitter-follower.

Another approach which has the additional appeal of simplicity of circuit construction, is to use negative feedback to increase the input impedance of an amplifier. The circuits of Figures 12 and 13 illustrate a two-stage DC amplifier which uses negative feedback to provide high input and low output impedance. The circuit was designed to have a 50-db power gain using the CA3018A integrated circuit. Calculations of the input impedance and power gain biased on the circuit of Figure 13 are presented in Appendix A.

As metal-oxide-semiconductor (MOS) techniques become more applicable to monolithic circuit integration techniques, the use of field-effect transistors (FET) will eliminate the need for Darlington pairs or feedback schemes to provide the required high input impedance. The FET has the additional, highly desirable characteristic of having the lowest noise of any device available for such use. At present FET's are available as input devices only in hybrid form - separately made then bonded to the IC chip. These hybrid circuits are quite expensive and of limited availability.

It is noteworthy that since tuned circuits cannot be integrated economically with existing technology, using integrated circuits for the other parts of the circuits is feasible only if a majority of the non-active components can be replaced by integrated active elements, or if greatly improved performance or stability can be obtained. Thus one finds the "diode stack" configuration of transistors used as a voltage divider or current source, the diode configured transistor used as an internal temperature-compensation element and other active element uses that are justifiable because of the ease of manufacture and the closely matched characteristics obtainable with monolithic transistor circuits.

Thus the circuits illustrated in Figures 12 and 13 are not as attractive to the designer as that of Figure 3 because of the many external non-active components.

An inspection of Figure 3 shows that the bias of Q4 is determined by Q1. Note that the base of Q4 is at the same voltage as the base and emitter of Q1. Therefore, since Q1 and Q4 are identical (or nearly so) their collector currents will be equal regardless of the collector load on Q4. Thus Q1 functions as an ideal voltage source. It is just this property that, while excellent for biasing, causes the very low input impedance to the amplifier and causes the tank circuit loading problems.

A typical integrated circuit which uses the diode-connected transistor for biasing is the Fairchild Semiconductor μA703 shown schematically in Figure 14. An RF amplifier based upon the μA703 is shown in Figure 15. Note that there are no external biasing elements. It should also be noted that the input resistance is

1.7K and that for stability and maximum power transfer, the input and output must be conjugately matched to the device. Table II lists the parameters of the μA703 connected as a 30MHz amplifier. [Reference 2].

It would appear from the considerations of gain, stability and input impedance that a solution would be to integrate the resistors required in the circuit of Figure 13, with the exception of R_B which is large enough (25K) to cause difficulty in construction. Should AGC be required with circuits shown in Figures 12 and 13, the biasing technique of using R_B between the emitter of Q4 and the base of Q3 would severely limit AGC action. This method of biasing provides gain stability for variations in device parameter. This can be understood as follows: Suppose the emitter current of Q4 decreases, the base current of Q3 thru R_B will then go down causing the collector voltage of Q3 to rise and thus drive Q4 harder. The net effect is therefore a cancellation of AGC action.

The Motorola MC1550 is an integrated circuit in which the biasing resistors are integrated leaving only tuned circuits and supply as external connections. A schematic diagram of the MC1550 connected as a 30-MHz tuned amplifier is shown in Figure 15. The MC1515 is also suitable to be configured as a mixer: Note the LC's similarity to one half of Figure 5 [Reference 1].

A schematic diagram of a suggested complete "receiver card" is shown in Figure 15. Although the integrated circuits used are of different manufacture, they are compatible insofar as signal levels and power supply voltages are concerned.

FIGURE 14. 30-MHz RF amplifier using Fairchild semiconductor μA703.

	T1	T2
Pri	10T	12T
Sec	10T	1T

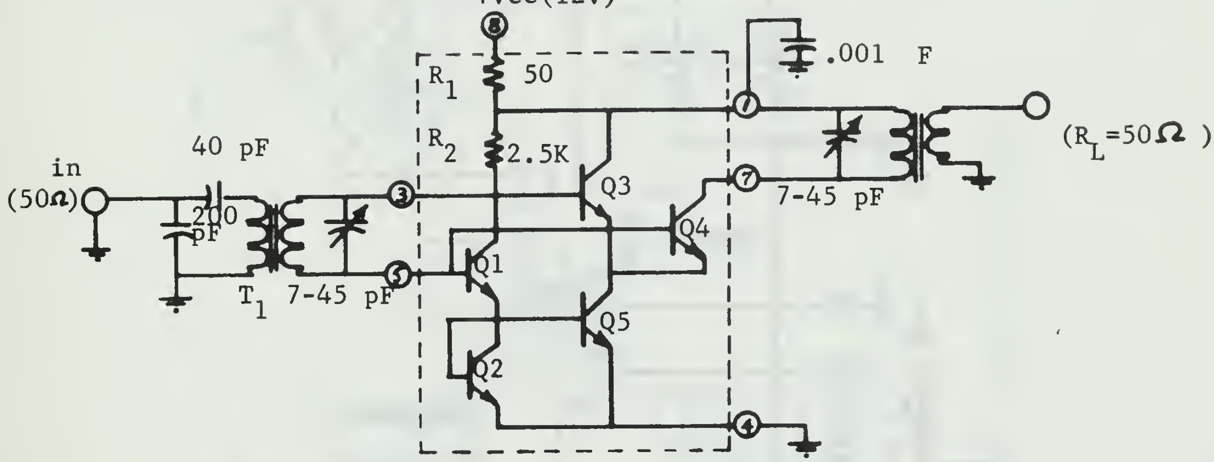
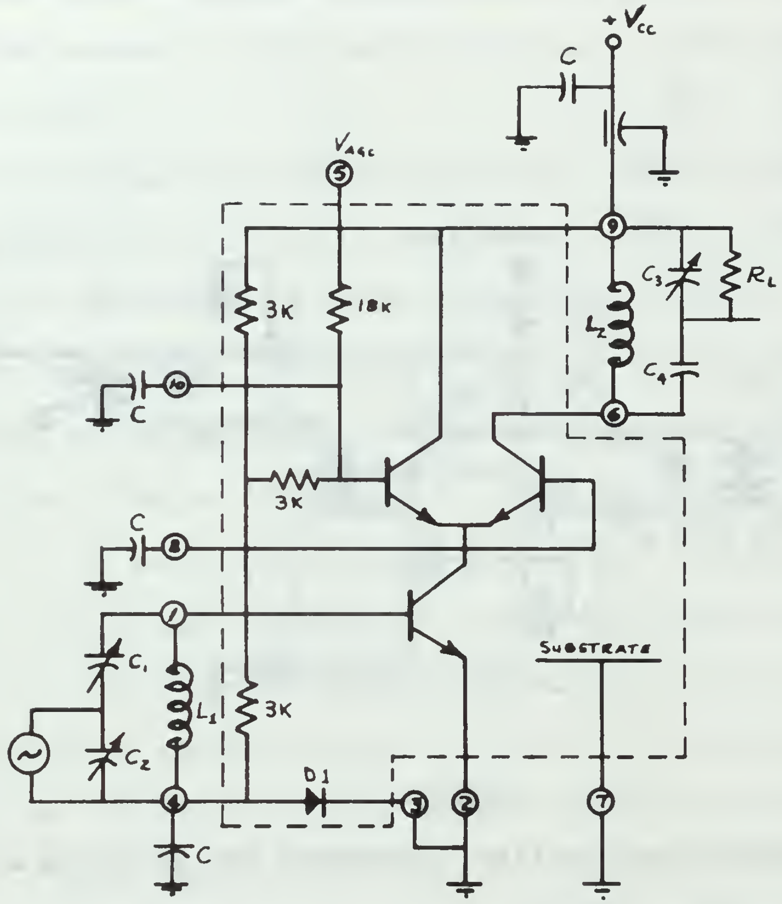


TABLE II

Device Parameters and Amplifier Performance For μA703 as a 30-MHz RF Amplifier (Ref. 2).

Input Resistance	1.7 K
Input Capacitance	9.0 pF
Output Conductance	0.08 mmho
Output Capacitance	2.5 pF
Forward Transadmittance	32 $\angle 140°$ mmho
Reverse Transadmittance	0.004 mmho
Power Gain	35 dB
Bandwidth	1 MHz
Noise Figure	6 dB
Maximum Stable Gain	39 dB



L_1 = 0.92 H

C_1 = 48 pF

C_2 = 88 pF

C = 2000 pF (bypass)

L_2 = 1.04 H

C_3 = 740 pF

C_4 = 33 pF

R_L = 50

FIGURE 15. 30-MHz tuned amplifier using Motorola MC1550 Integrated Circuit

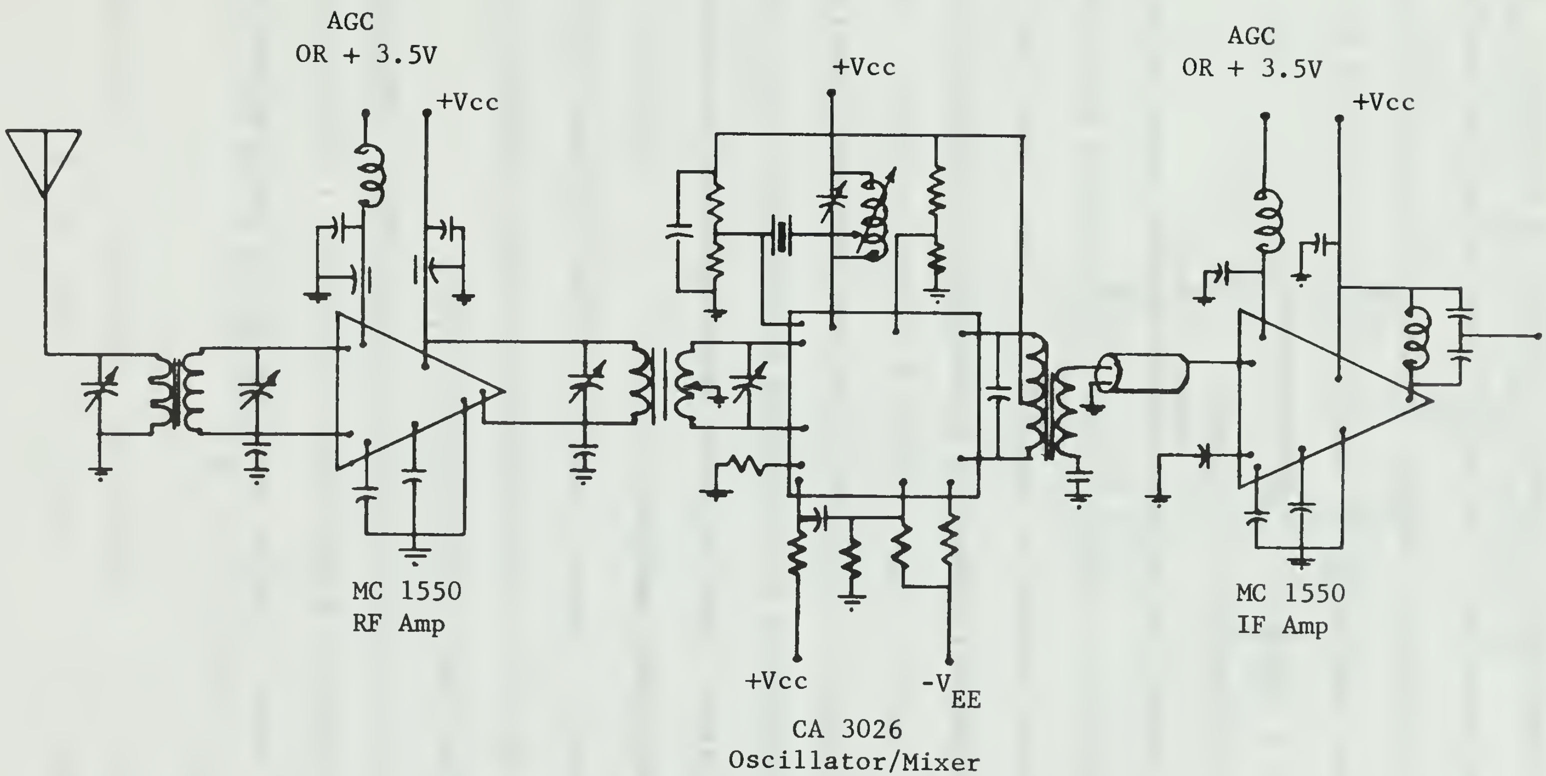


FIGURE 16. Channelized receiver using integrated circuits

A. RECOMMENDATIONS FOR FURTHER WORK

The block diagram of Figure 1 shows the overall system for which the channelized, fixed-tuned receiver is envisioned. With the possible exception of the audio output stages, all sections of the receiver system are amenable to monolithic integrated-circuit construction. Most of the integrated circuits presented here are suitable for a wide variety of uses and hence could be used throughout a system such as Figure 1.

Since coverage of the entire HF RF spectrum might be desired, the large number of "receiver cards" required makes manual switching of the receivers between an appropriate antenna, power supply and the wideband amplifier stage an impossibility. Electronic switching is the only viable means of channel selection. Switching of the power supply connections is desirable because it allows a significant reduction in the standby drain current. Only that receiver section in use need have power supplied to it since there is no "warm-up" time involved.

An operator's console containing display devices, signal-processing equipment and an electronic, digital channel-selection panel is envisioned to complete the system.

The electronic digital-switching system might prove to be a worthwhile topic for further thesis work since it would involve aspects of control, digital equipment design and transmission-line theory.

APPENDIX A

Calculations for amplifier shown in Figure 13.

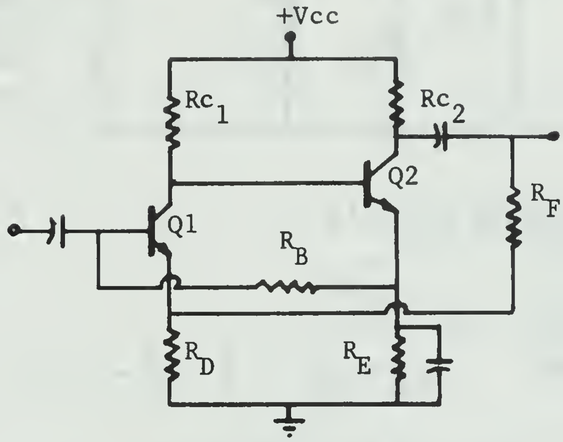


FIGURE 17. Feedback Amplifier. Data taken from RCA CA3018A.

$Ic_1 = 1mA$, $Ic_3 = 3mA$

A voltage gain of 100 is desired. Therefore since $Av \leq R_F/R_D$, make $R_F/R_D = 110$. Choosing R_D small so as to make the open-loop gain as high as possible, let $R_D = 50\Omega$, then $R_F = 5.6K$.

For a 6-V supply, $V_{c_2} = 3v$ for optimum bias. Selecting $I_{c_2} = 3mA$ and $I_{c_1} = 1mA$, $R_{c_2} = 1K\Omega$.

Let the emitter voltage of Q2 = 1v, then

$$R_E = 1.0v/3.03mA = 330\Omega$$

then since $V_{B_1} = 0.75v$ and $I_{B_1} = 0.01\ R_B$ is

$$R_B = 0.25v/0.01mA = 25K\Omega.$$

$V_{c_1} = 1.7v$ as set by V_{B_2} therefore

$$R_{c_1} = 4.3v/1.03mA = 4.17K\Omega.$$

Summarizing: $R_{c_1} = 4.17K\Omega$ $R_E = 330\Omega$ $R_F = 5.6K\Omega$

$R_{c_2} = 1.0K\Omega$ $R_D = 50\Omega$ $R_B = 25K\Omega$

The open-loop AC circuit can be modeled as shown in Figure 18

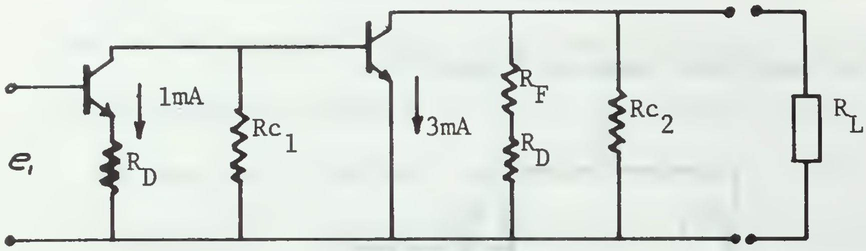


FIGURE 18
Simplified AC Model of Feedback Amplifier

The gain of the second stage is

$$A_2 = \frac{-g_{m2}}{\Sigma g_L}$$

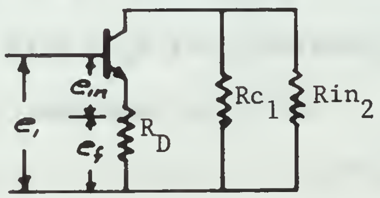
$g_{m2} = I_c \times 38$ mmho $\qquad \sum g_L = \frac{1}{5.6} + \frac{1}{1} + \frac{1}{R_L} = 1.18.$

Assuming R_L to be laige and ignoring Ro_2.

$$A_2 = -\frac{3 \times 38}{1.18} = 96.5.$$

Now $R_{in2} = \frac{h_{fe}}{g_{m2}} = \frac{100}{114} = .88K.$

Then the first stage can be modeled as shown below.



The feedback (degenerative) fraction,
$F = Rd/R_{in2} = \frac{50}{880}$
$= 0.057.$

The open-loop gain, A_1 is therefore

$$A_1 = -gm_1 \;\; RL = -\ (38 \times 1)(-930) = 35.$$

Then with degenerative feedback included:

$$A_{1f} = \frac{A_1}{1 + A_1F_1} = \frac{-35}{1 + 35(.057)} = 11.7.$$

The total open-loop gain of both stage is therefore

$$A_T = 96.5 \times 11.7 = 1130.$$

The total gain with feedback is

$$A_F = \frac{A_T}{1 + FA_T} = \frac{1130}{1 + (.0091)(1130)}$$

where $F = R_D/R_F = 0.0091$

$$\text{and} \quad A_F = \frac{1130}{11.3} = 100.$$

The input resistance to the first stage, without overall feedback is

$$R_{in1} = \frac{e_1}{i_1} = (1 + A_1F_1)\left(\frac{h_{fe1}}{gin_1}\right)$$

$$= (1 + 35 \times .057)\left(\frac{110}{38}\right)$$

$$= 8.7K\Omega$$

Then R_{in}, taking feedback into account is

$$R_{in} = (1 + FA_T)(R_{in1})$$

$$= (11.3)\ (8.7K)$$

$$= \underline{98K}\Omega$$

The current gain is

$$A_i = A_v \frac{R_{in}}{R_L} \quad .$$

Assuming a 10K load at the resonant frequency of the input tank of the mixer.

$$A_i = \frac{100 \times 98}{10K} = 980.$$

The power gain is

$$G = A_vA_i = 100 \times 980 = 0.8 \times 10^4$$

$$\text{or} \quad G_{db} = 10 \log_{10}(G) = \underline{49.9 \text{ dB}} \ .$$

IMPROVED OSCILLATOR CIRCUIT

The circuit shown in Figure 19 was found to perform better than that of Figure 6 using the same crystal.

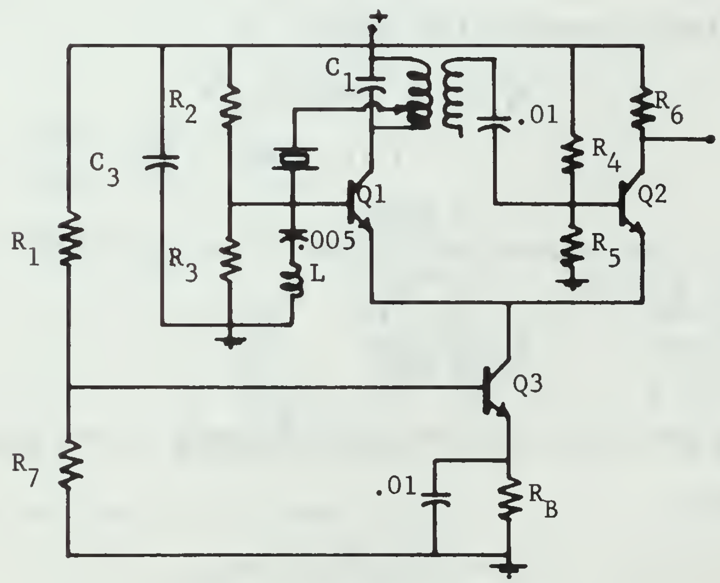


FIGURE 19. Modified 30-MHz crystal-controlled oscillator using RCA CA3018A I.C.

APPENDIX B

PHOTOGRAPHS

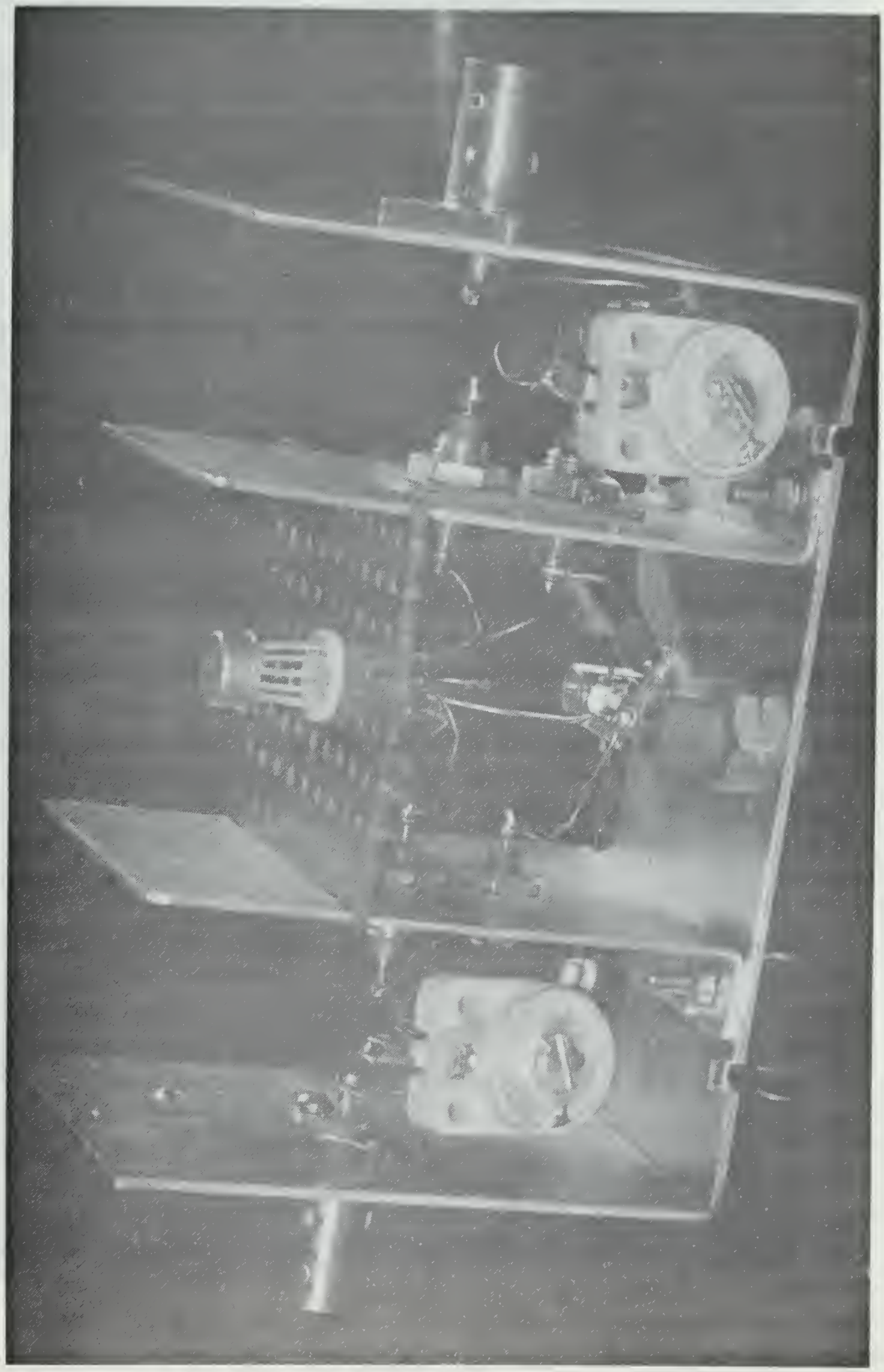


FIGURE 20. RF AMPLIFIER

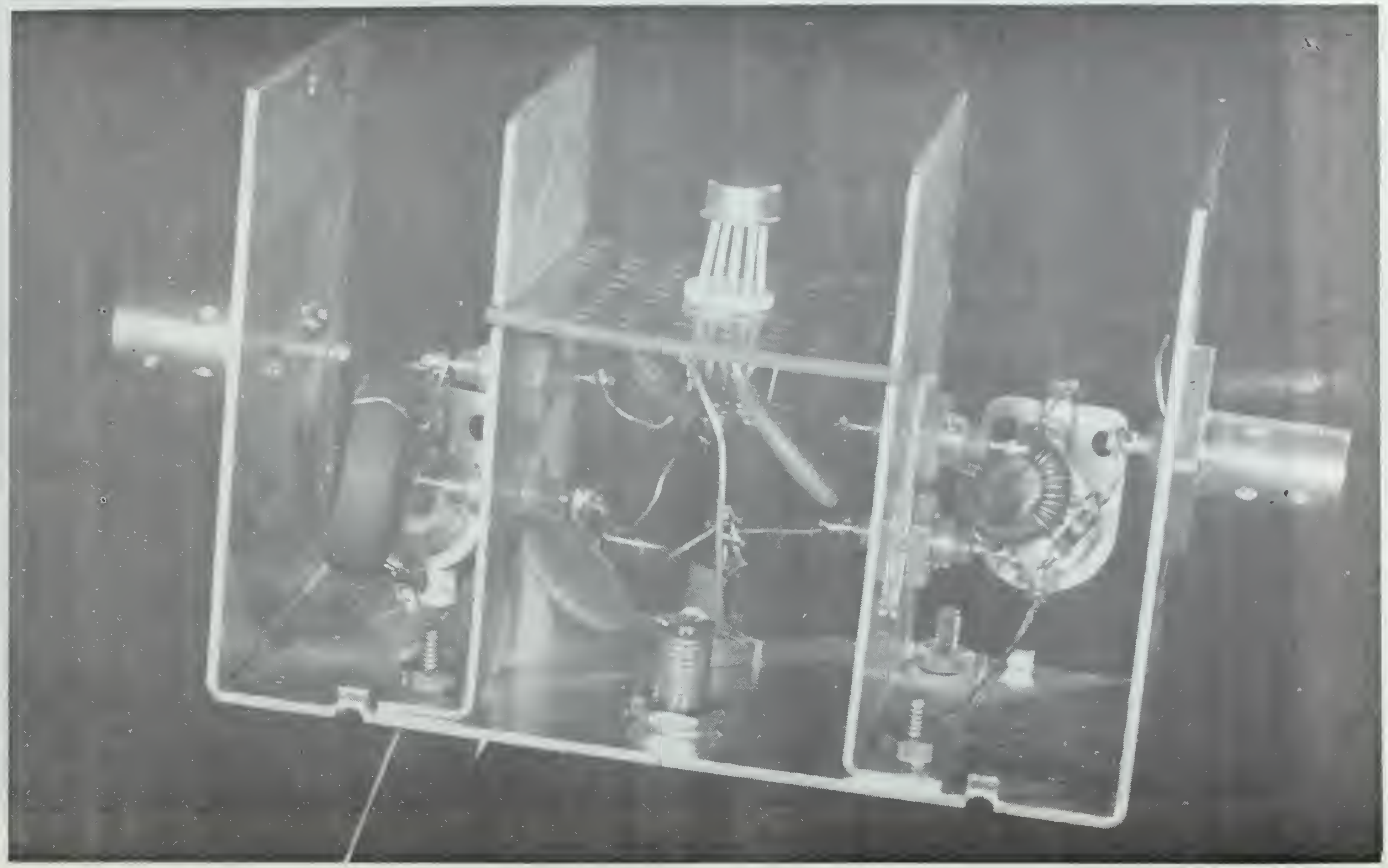


FIGURE 21. RF AMPLIFIER

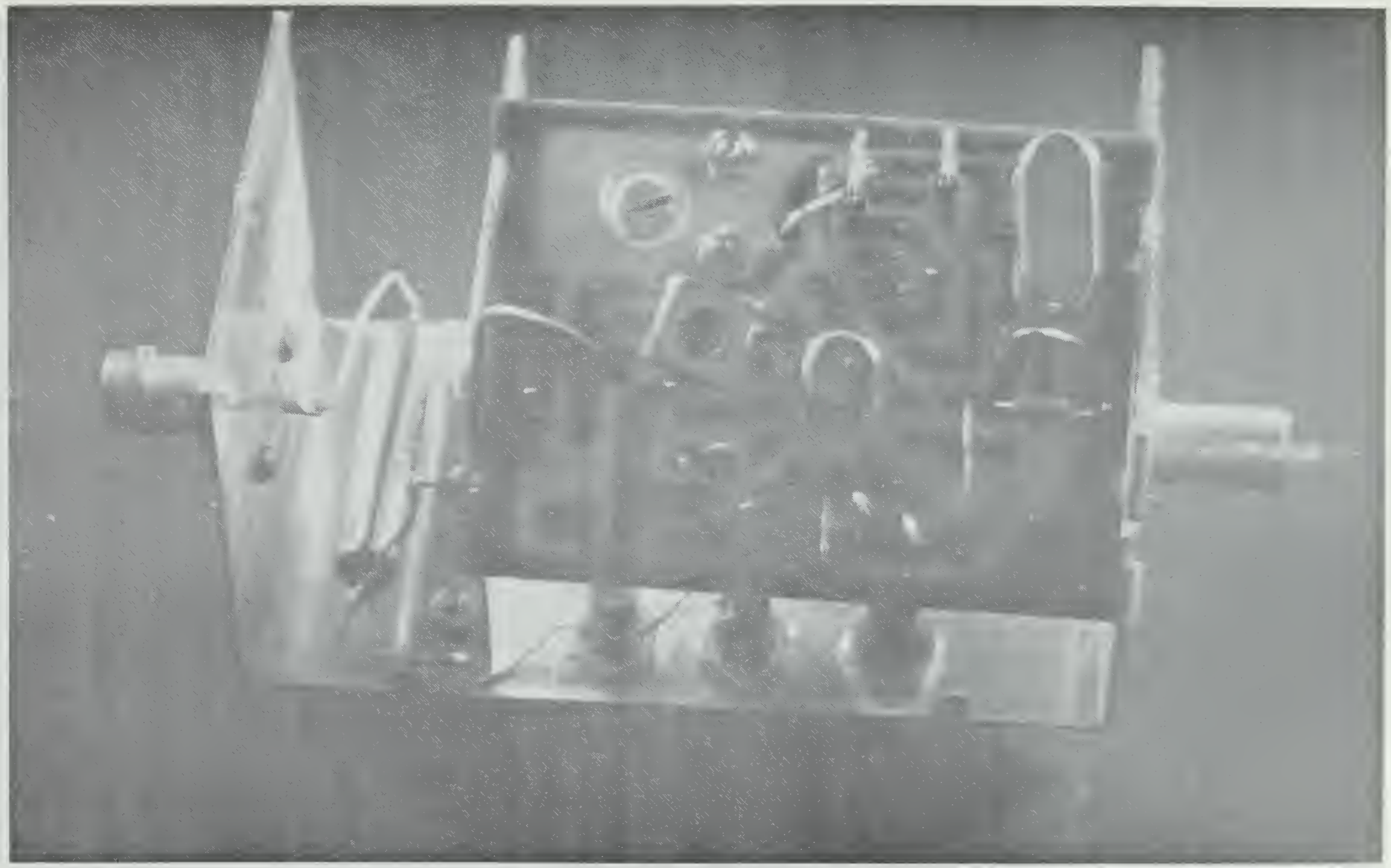


FIGURE 22. FREQUENCY CONVERSION STAGE. BACK LIGHTED TO SHOW PC BOARD

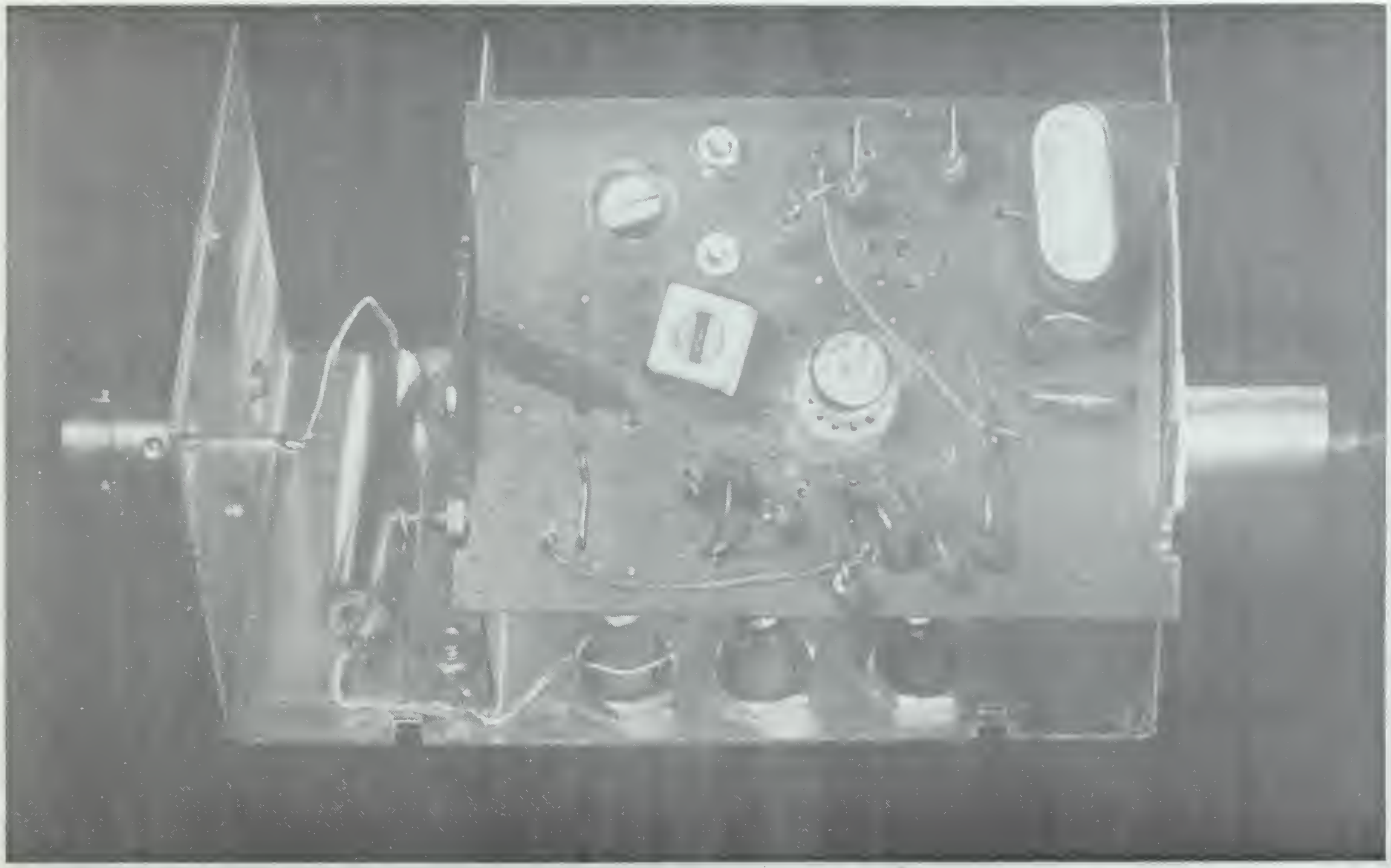


FIGURE 23. FREQUENCY CONVERSION STAGE

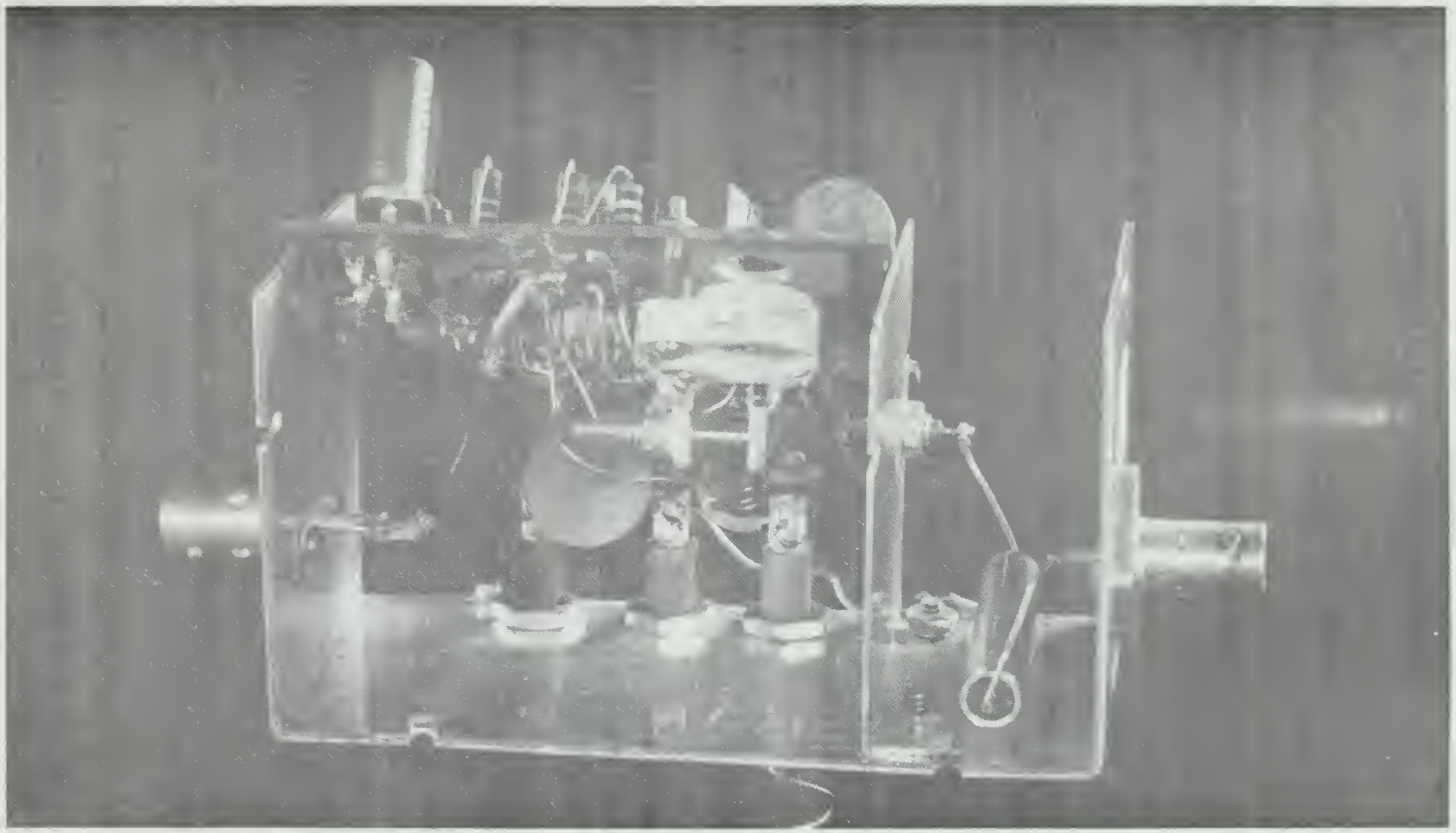


FIGURE 24. FREQUENCY CONVERSION STAGE

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13. ABSTRACT

An application of modern linear integrated circuits to the design of a channelized high-frequency radio receiver is presented. The results of testing a prototype, crystal-controlled, fixed-tuned integrated-circuit receiver front end and wideband frequency-conversion stage are shown. Problems of image rejection with single frequency conversion are discussed. The design of a channelized receiver section using modern special-purpose linear integrated circuits is shown.

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